ASSIGNMENT
Of

EXPERIMENTAL TECHNIQUES IN NUCLEAR AND PARTICLE PHYSICS

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Q1. Explain delay line wave shaping of detector pulse. How do we get bipolar pulse with this technique? (8)

Ans. 1. <u>SINGLE DELAY LINE (SDL) SHAPING</u>

The properties of coaxial cables can be used in pulse shaping. We know that a coaxial cable that is shorted at the receiving end will give rise to a reflection when a step voltage reaches that end of the cable. The reflection is a step moving back toward the sensing end of the cable, with an amplitude equal to the initial step but opposite in polarity.
A configuration in which a shorted transmission line can be used to shape a step input is shown in the fig. 1. The transmission line or coaxial cable is assumed to be long enough so that the propagation time through its entire length is long compared with the rise time of the step voltage applied to the input of the network. The unity-gain operational amplifier simply provides impedance isolation at the input. Because the amplifier is assumed to have zero output impedance, the resistor Z_0 terminates the cable in its own characteristic impedance at the sending end. The far end of the cable is assumed to be shorted, or terminated in zero resistance.

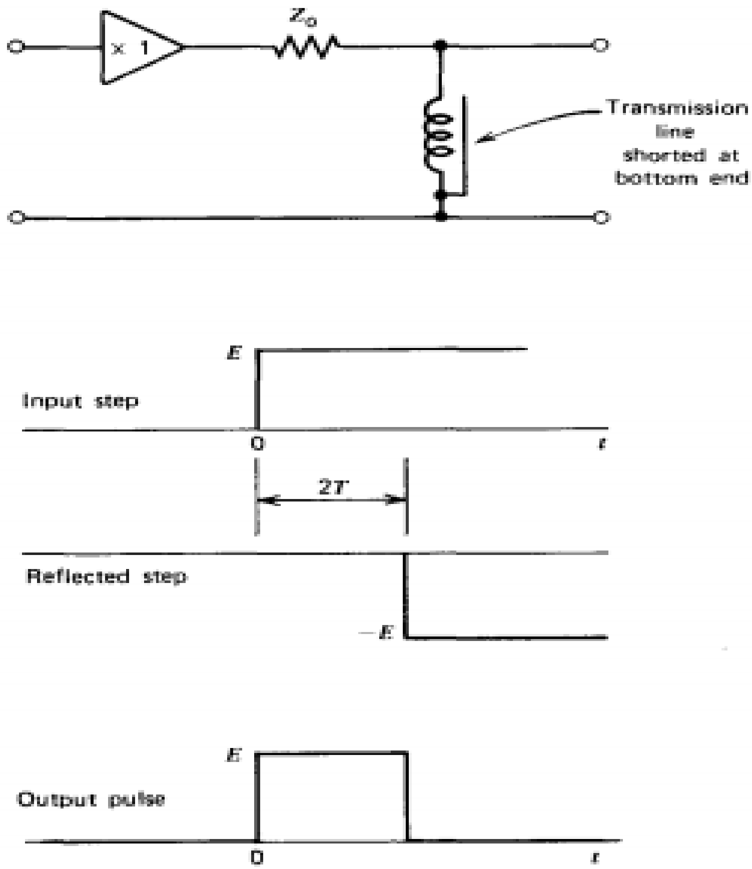


Figure 1 Single Delay Line (SDL) network applied to shaping an input step waveform

The waveforms shown in fig.1 illustrate the sequence of events if a step voltage is applied to the input of the network. The initial positive step is assumed to be applied at t = 0 and to persist for a long period of time. The step propagates down the cable, is reflected form the shorted end, and travels back to the sending end

as inverted or negative step, where further reflections are prevented by the impedance matching resistor Z_0. The voltage observed at the output of the network is simply the algebraic sum of the original and reflected waveforms, or the rectangular pulse shown in the figure. The time width of this shaped pulse is the down-and-back propagation time through the length of cable used for shaping. For many applications, this propagation time should be of the order of a microsecond. In order to avoid excessively long cables, special delay lines with much reduced propagation velocity are often used for this purpose. Hence the process described is often called **single delay line (SDL) shaping.**

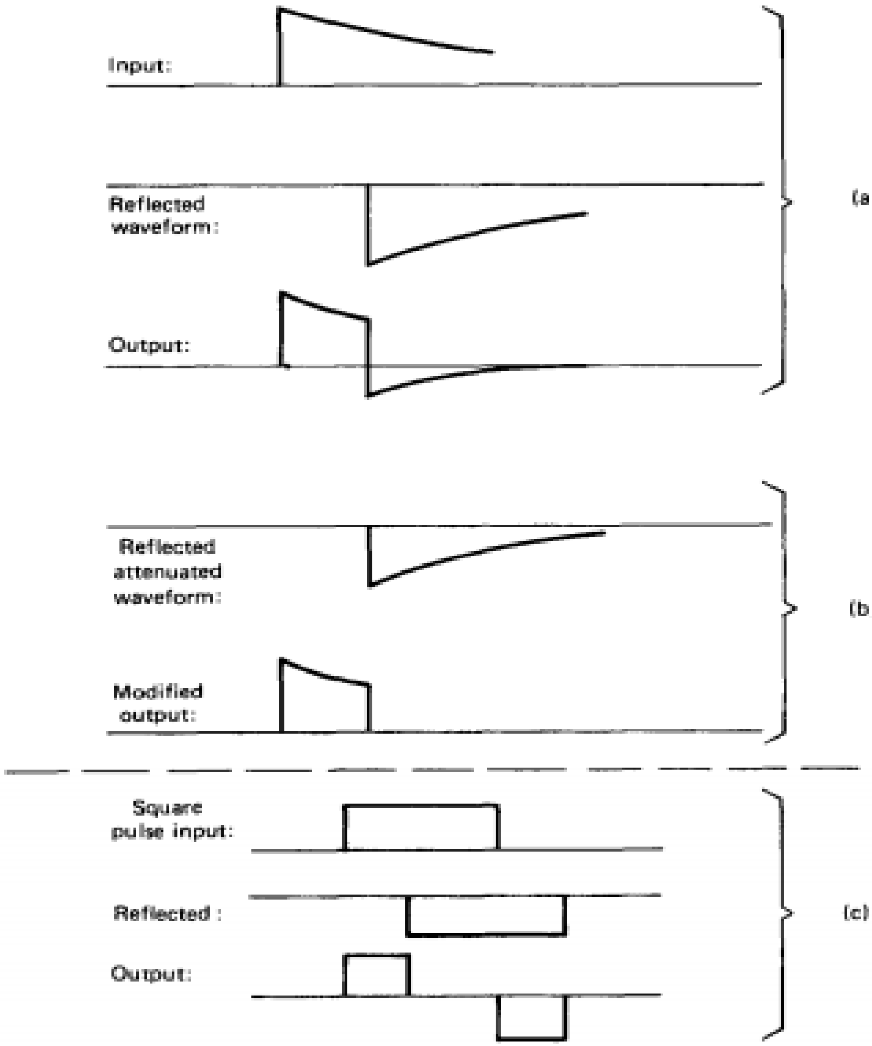


Figure 2 (a) Effect of applying simple SDL shaping to an input tail pulse with decay time comparable with the propagation time of delay line. (b) Remedy to the undershoot of part (a) accomplished by partially attenuating the reflected waveform. (c) Effect of SDL shaping on a rectangular shaped input pulse whose length exceeds the delay line down-and-back time.

The propagation time of the delay line must always be larger than the rise time of the pulse from the preamplifier to avoid the ballistic deficit in RC shaping. Also, if the decay time of the preamplifier pulse is not many times larger than the propagation time, a situation will result as illustrated in fig. 2(a), in which there is undesirable undershoot following the shaped pulse. If the preamplifier decay time is always a fixed value, this undershoot can virtually be eliminated by slightly

attenuating the reflected pulse as shown. The attenuation can be accomplished by raising the termination resistance at the reflecting end of the cable from zero to a small value that is less than the cable characteristic impedance. Similar shaping is also used to reduce the length of pulses with very fast (≈ns) leading edges. In this case the role of the delay line is carried out by a normal coaxial cable of a few meters length, commonly called a *shorted stub.* As is shown in fig. 2(c), this type of shaping applied to a rectangular output input pulse gives rise to two shaped pulses of opposite polarity separated in time by the length of the original rectangular pulse.

2. DOUBLE DELAY LINE (DDL) SHAPING

Because the output of an SDL shaping network is a unipolar pulse for typical preamplifier pulses, the same comments apply as were made earlier for single differentiating networks. At high counting rates, it may be desirable to substitute a shaping network that results in *bipolar pulses* to eliminate baseline shift. **Bipolar pulses** can be produced if the *two delay lines* are used in the configuration shown in Fig.3, called **double delay line (DDL) shaping.** Both delay lines should have equal propagation time, and the resulting impulse will have positive and negative lobes of equal amplitude and duration. Therefore, an average dc level of zero can accurately be maintained and baseline shift in subsequent ac coupled circuits will virtually be eliminated. Although DDL shaping has high counting rate characteristics, it applies no high-frequency filtering to the signal and is thus usually inferior from a signal-to-noise stand point compared with methods based on RC circuits.

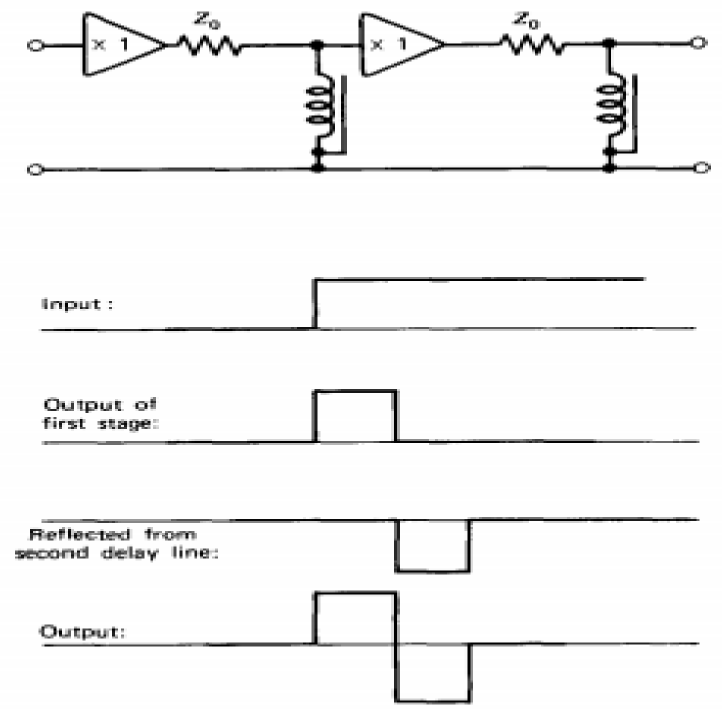


Figure 3 A DDL shaping network with equal delay times applied to a step input waveform

Q2. Discuss the charge-sensitive configuration of pre-amplifier and derive the expression for V_{out}. (8)

Ans. <u>Preamplifier</u> : The first element in a signal-processing chain is therefore often a preamplifier (or "preamp") provided as an interface between the detector and the pulse-processing and analysis electronics that follow. The preamplifier is usually located as close as possible to the detector. From a signal to-noise standpoint, it is always preferable to minimize the capacitive loading on the preamp, and therefore long interconnecting cables between the detector and preamp should be avoided if possible. One function of the preamp is to terminate the capacitance quickly and therefore to maximize the signal-to-noise ratio. Because of convenience or safety considerations, the components that follow in the pulse processing chain often are located at some distance from the detector and preamp. Thus another requirement is for the preamp output stage to be capable of driving its signal into the large capacitance represented by the long interconnecting cable, or that it have a low output impedance.

The preamplifier conventionally provides no pulse shaping, and its output is a linear tail pulse. The rise time of the output pulse is kept as short as possible, consistent with the charge collection time in the detector itself. The decay time of the pulse is made quite large (typically 50 or 100 ps) so that full collection of the charge from detectors with widely differing collection times can occur before significant decay of the pulse sets in.

<u>Charge Sensitive Preamplifier</u> : Charge Sensitive Preamplifier is an electronic device that can integrate a current signal and generate a voltage signal with an amplitude proportional to the incoming input charge. The signal produced by many detectors is of very low amplitude and therefore needs adequate amplification. If the output signal can be assimilated to a current pulse then it is convenient to use a preamplifier that is based on a **charge sensitive preamplifier type (CSP)**. The current pulse generated by the detector is converted into a voltage pulse by means of the charge of a capacitor. The basic diagram of a charge sensitive pre-amplifier is shown in the fig.4.

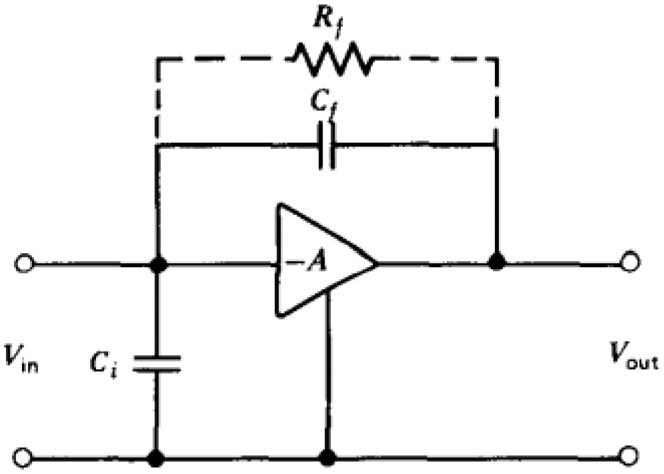


Figure 4. Simplified diagram of a charge-sensitive preamplifier configuration

The elements of a charge-sensitive configuration that can remedy this situation are shown in Fig. 4. For this circuit, the output voltage is proportional to the total integrated charge in the pulse provided to the input terminals, as long as the duration of the input pulse is short compared with the time constant R_fC_f Changes in the input capacitance no longer have an appreciable effect on the output voltage. Although originally developed for use with semiconductor diode detectors, this charge-sensitive configuration has proved its superiority in a number of other applications, so that preamplifiers used with other detectors in which the capacitance does not necessarily change are also often of the charge-sensitive design.
Ideally, the rise time of the pulse produced from the preamplifier is determined only by the charge collection time in the detector and is independent of the capacitance of the detector or preamp input. That is generally the case, but for fast detectors the rise time may also be influenced by time constants arising from several factors. Prominent among these is any series resistance from undepleted regions in semiconductor detectors or imperfect electrical contact to the detector active volume. This resistance couples with capacitance at the input of the preamplifier to define a time constant that can slow the rise of the pulse if it is comparable with or exceeds the charge collection time.

EXPRESSION FOR V_{OUT} :

Assume $A >> (C_i + C_f)/C_f$

$$V_{out} = -A\ V_{in}$$

$$V_{out} = -A \frac{Q}{C_i + (A+1)C_f}$$

$$V_{out} \cong -\frac{Q}{C_f}$$

Where Q is the charge collected on the detector.

Q3. Define Walk and Jitter. Explain the various time pick-off methods for time measurement. (12)

Ans. **<u>Time Pick-Off Methods</u>**

The most fundamental operation in time measurement is the generation of a logic pulse whose leading edge indicates the time of occurrence of an input linear pulse. Electronic devices that carry out this function are called *time pick-off* units or *triggers.*
Factors that lead to some degree of uncertainty in deriving the timing signal are always present. Sources of timing inaccuracy are divided into two categories.

Those that apply when the input pulse amplitude is constant are usually called sources of time jitter, whereas those effects that derive primarily from the variable amplitudes of input pulses are grouped together in a category called amplitude walk or time slewing. Relative importance of these two categories depends on the dynamic range expected in the input pulse amplitude. The best timing performance will be achieved if the input pulses are confined to a very narrow range in amplitude, because then only sources of time jitter contribute to uncertainty. However, practical applications often require that pulses of different amplitudes be processed, and the additional contribution of walk will worsen the overall time resolution of the system.

1. LEADING EDGE TRIGGERING.

The most direct method is to sense the time that the pulse crosses a fixed discrimination level. Such **leading-edge** timing methods are in common use and can be quite effective, especially in situations in which the dynamic ranges of the input pulses are not large.

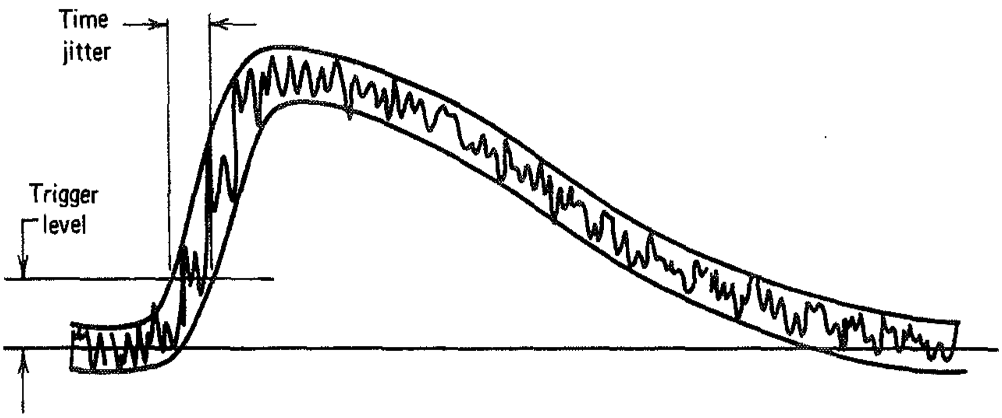


Figure 5. The time jitter in leading edge triggering arising from random noise. An envelope is shown of many repeated signal pulses of the same amplitude and shape, but with a random contribution of noise.

The effect of time jitter on leading edge timing is shown in the fig.5. The random fluctuations superimposed on signal pulses of identical size and shape may cause the generation of an output logic pulse at somewhat different times w.r.t. the centroid of the pulse. The timing errors will be approximately symmetrical and will increase if the slope of the leading edge of the pulse is decreased.

The amplitude walk associated with a leading-edge trigger is graphically demonstrated in fig. 6. The two pulses shown have identical true time of origin but give rise to output logic pulses that differ substantially in their timing. Under extreme conditions, the amplitude walk can amount to the full ruse time of the input pulse and is often unacceptable in situations where accurate timing is needed over a wide amplitude range.

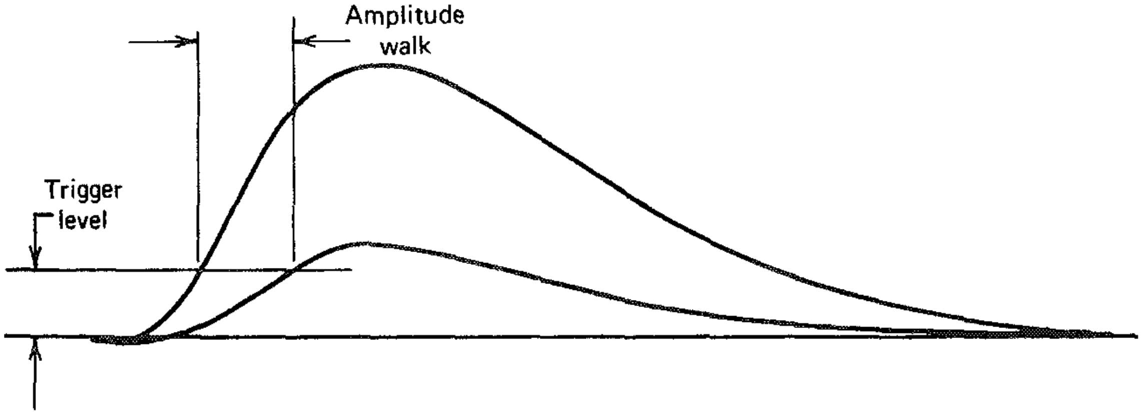


Figure 6. Amplitude walk in leading edge triggering. Two pulses with identical shape and time of occurrence but different amplitude are seen to cross the trigger level at different times.

Even if the amplitude is constant, walk can still take place if changes occur in the shape of the pulse. Detectors with variable charge collection time, such as germanium detectors, produce output pulses with variable rise time as shown in fig. 7. Change that occur in the pulse shape before the discrimination point will affect the timing and can constitute another source of timing uncertainty. The sensitivity of leading-edge triggering to timing walk because of amplitude and shape variations is minimized by setting the discrimination level as low as possible. However, the discrimination point should be in a region of steep slope on the pulse leading edge to minimize uncertainties due to jitter. Practical compromises in these conflicting often lead to optimum resolution for levels that are set at about 10-20% of the average pulse amplitude.

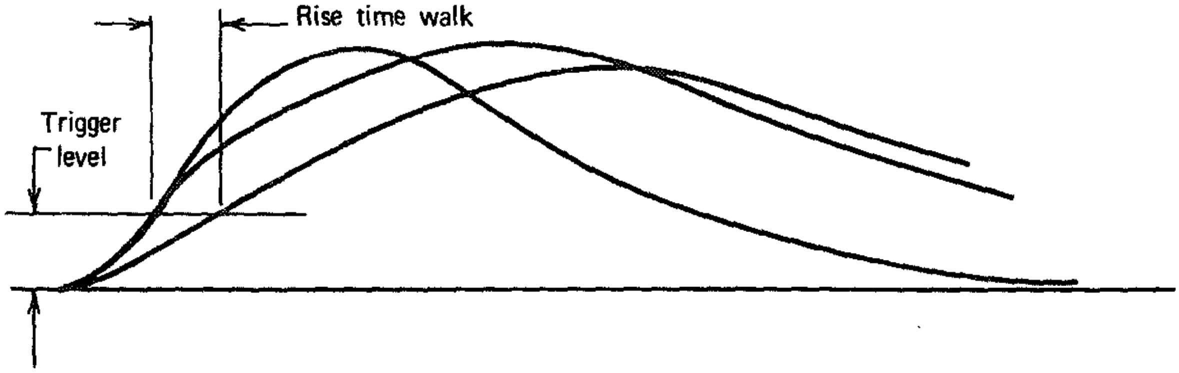


Figure 7. Time walk in leading edge triggering due to variations in pulse shape or rise time. Changes in either the overall pulse rise time or the shape of the leading edge for pulses with the same rise time can be responsible for time walk.

2. CROSSOVER TIMING.

When leading edge triggering is applied to pulses with a wide amplitude range, amplitude walk often results in large timing uncertainties. The crossover timing method can greatly reduce magnitude of amplitude walk but requires the input

pulse to have bipolar shape. Fig.8 shows two pulses with different amplitudes as shaped by a CR-RC-CR double differentiating network. Although the pulse amplitudes differ widely, the time at which the waveform crosses from the positive to the negative side of the axis is theoretically independent of the amplitude and depends only on the shaping time constant chosen for the network. Double Delay line shaping produces the same effect and is usually preferred in crossover timing. Compared with leading edge timing, crossover methods greatly reduce amplitude walk, but only at the expense of increased jitter. The noise introduced by the required shaping stage and the increased susceptibility of the crossover point to statistical fluctuations in the signal contribute to the increased jitter.

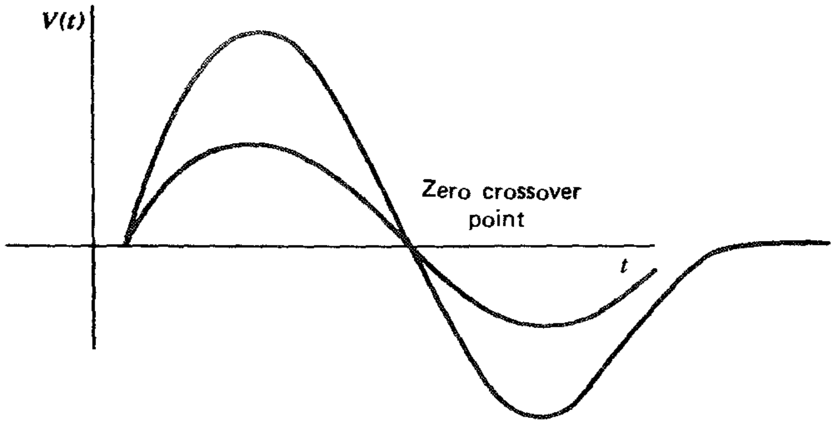


Figure 8. Bipolar pulses of different amplitude showing the same time of zero crossover.

A related method called *fast crossover* can be implemented for the specific case of fast scintillator-photomultiplier tube signals. A bipolar pulse is produced directly at the PMT anode using “tee” connection and a shorted length of coaxial cable. Because the time to the zero-crossover point is determined by the length of the shorted cable, it is also independent of the signal pulse amplitude. However, the pulse shape must not vary if this freedom from walk is to be preserved.

3. CONSTANT FRACTION TIMING.

If the input dynamic range is small, the lower jitter of leading edge timing results in superior timing performance compared with crossover timing. Furthermore, it is empirically found that the best leading edge timing characteristics are obtained when the timing discriminator is set at about 10-20% of the pulse amplitude. These observations have led the development of a time pick-off method that produces an output signal a fixed time after the leading edge of the pulse has reached a *constant fraction* of the peak pulse amplitude. This point is then independent of pulse amplitude for all pulses of constant shape. Therefore, pulses over a wide dynamic range can be accepted with much the same freedom from amplitude walk as exhibited by crossover timing, but with lower jitter. The electronic shaping steps required to carry out constant fraction timing are diagrammed in fig.9.

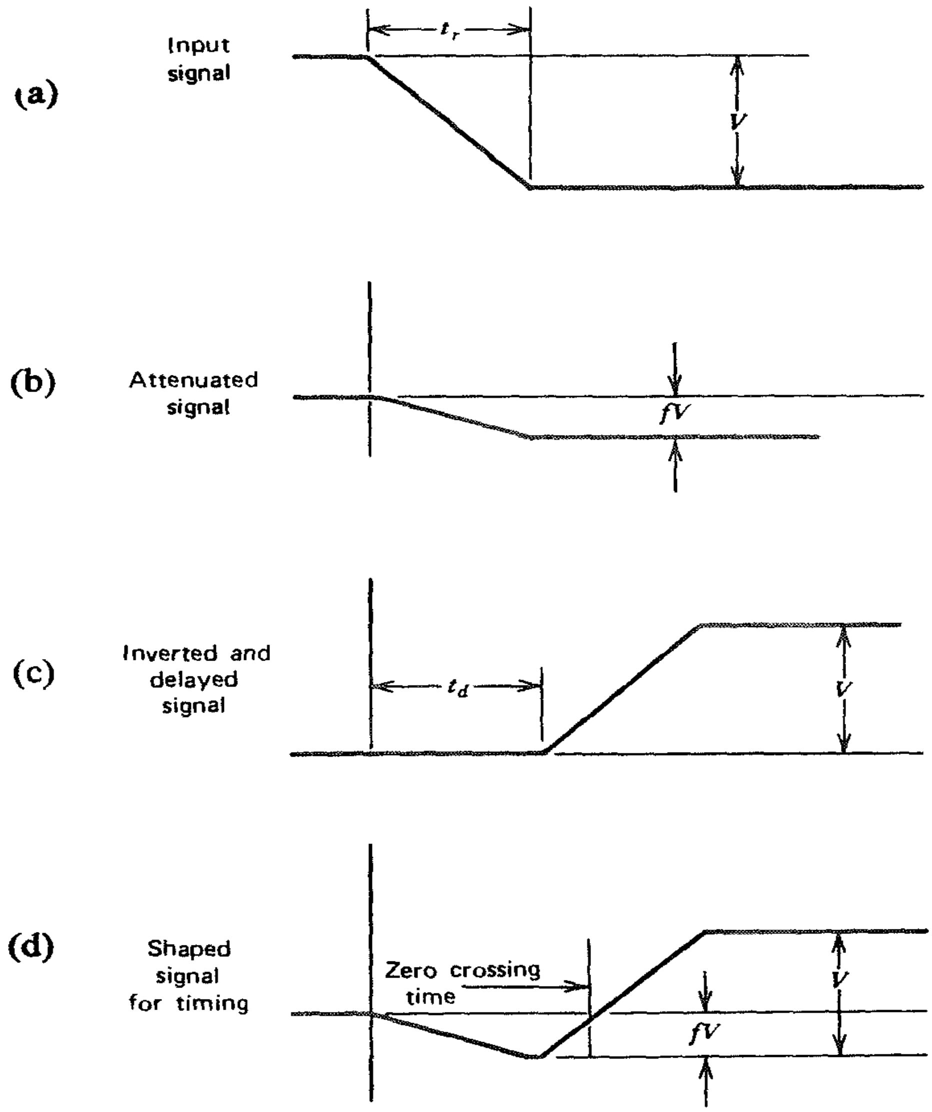


Figure 9. Waveforms in the constant fraction time pick-off method. For clarity, only the leading edge of the pulse is shown.

The process involves taking the preamplifier output [part(a)] and multiplying it by the fraction f that is to correspond to the desired timing fraction of full amplitude. The input waveform is also inverted and delayed for a time greater than the pulse rise time to give the waveform show in part (c). The sum of waveforms (b) and (c) is then taken to give the pulse that is shown in part (d). The time that this pulse crosses the zero axis is independent of the pulse amplitude and corresponds to the time at which the pulse reaches the fraction f of its final amplitude.

4. AMPLITUDE AND RISE TIME COMPENSATED (ARC) TIMING.

In situations where the shape or rise time of the pulse can change, even constant fraction timing methods cannot eliminate walk. To meet the need for accurate timing for germanium detectors, in which the rise time variations are large, the ARC timing systems are used. The scheme amounts to basing the timing signal on a fixed fraction of only the early portion of the pulses and is therefore unaffected by shape changes that may occur at a late point in the waveform.

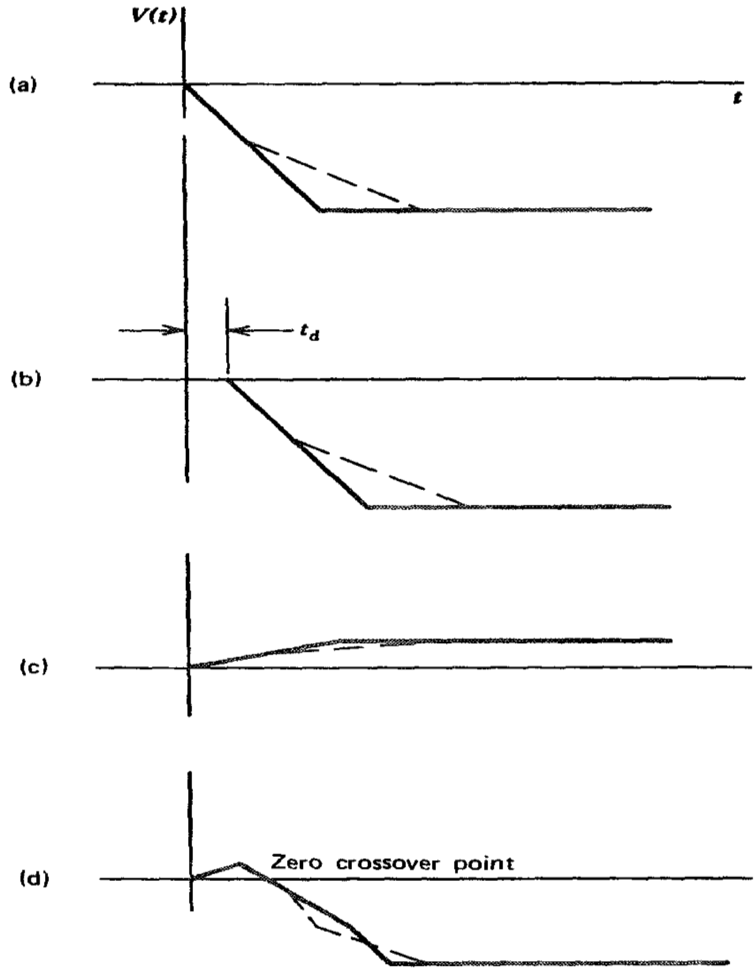


Figure 10. Waveforms in the ARC time pick-off method. The original waveform (*a*) is delayed (*b*), inverted, and attenuated (*c*). The sum of (*b*) and (*c*) gives waveform (*d*) with a zero crossover point used to generate the timing signal. The dashed waveform has a slower rise time than the solid, but both give the same timing signal.

The pulse-processing methods are similar to those used in constant fraction timing and are shown in the fig.10. The only difference is that the delay time illustrated in the figure is chosen to be as small as practical (≈5-10 ns). The delay must be sufficiently long so that the initial portion of the output pulse is allowed to significantly exceed the noise level, so that a simple discriminator signal can be generated that will actuate the circuits that sense the crossing of the waveform back across zero. Using this method, the time of zero crossing is given by $t_d/(1-f)$, provided that the input pulse rises linearly to that point. Therefore, this crossover time again occurs at a constant fraction of the amplitude of all input pulses and will furthermore be insensitive to any slope change in the rise of the input pulse that occurs after the zero crossing point.

5. EXTRAPOLATED LEADING EDGE (ELET) TIMING.

An alternative to ARC timing, an extrapolation principle has also been applied successfully to the almost walk-free derivation of timing signals from germanium detectors with variable rise time. The method is similar to ARC timing in that it is based only on the initial portion of the pulse rise and is not affected by shape

variations that may follow that initial portion. ELET timing uses two independent discriminators set at a different discrimination levels to carry out an extrapolation of the pulse waveform back to its time of origin.

For illustration, assume that the second discrimination level is set at twice the level of the first. For pulses with a linear rise, the time difference between the discrimination points should be equal to the time delay between the true start of the pulse and the time at which the first discriminator is crossed. The technique employs time-to-amplitude convertors to carry out an effective extrapolation back to zero, so that a timing pulse corresponding to the true time of origin of the input pulse is generated. The two discrimination levels are ideally set as low as possible, so that the assumption of linearity need be met for only a short portion of the leading edge of the pulse.

6. FIRST PHOTOELECTRON (FPET) TIMING.

A unique method for providing timing information from scintillation detectors can be applied when the conditions of the experimental limit the possible triggering time to a relatively small time interval. If this interval is small compared with the average time spacing between photomultiplier tube noise pulses, excellent timing can be carried out simply by sensing the arrival of the photomultiplier signal corresponding to a single photoelectron. This is, in effect, leading edge timing with a trigger level that is as low as physically possible. Because PMT noise would also trigger at this level, the probability of a noise pulse within the time range of interest must remain small. This condition has been made practical by low-noise bialkali PMTs in which noise rates may be as low as tens or hundreds per second. The FPET is advantageous for timing from scintillators when low-energy gamma rays are involved.

Q4. Write note on : (i) Overload recovery and pile-up (ii) Time to amplitude converters. (8)

Ans. (i) Overload Recovery and Pile-up

The charge-sensitive preamplifier is relatively susceptible to saturation when very large pulses are supplied to its input. The overload recovery properties of the preamplifier are therefore an important specification for those applications in which frequent large pulses might obscure the signal to be measured.

Because output of a preamplifier is a tail pulse with a rather long decay time, the pile-up of pulses is inevitable, except at very low signal rates. Although the effects of this pile-up are largely removed by subsequent shaping, one potential effect cannot be dealt with so easily. In fig.11, the piled-up output of a preamplifier at high rates is sketched. The average level of this waveform will increase with rate and may approach the limit of linear operation of the preamplifier. In that event,

some piled-up pulses may drive the preamplifier into saturation and thus will be seriously distorted. Choosing small values for the feedback resistor R_f will minimize this effect by ensuring rapid decay of the pulse, but at the expense of an increased noise contribution.

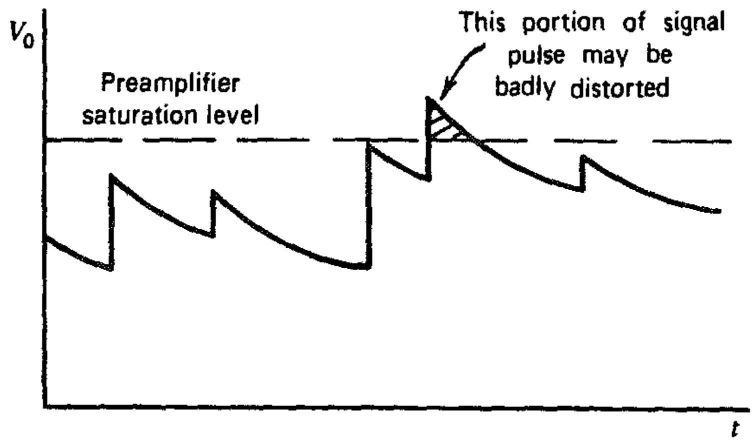


Figure 11. The pile-up of pulses within the preamplifier at high rates. If the saturation level of the preamplifier is exceeded, some pulses can be seriously distorted.

The pulse rate at which saturation occurs depends on the type of connection between the preamp and the detector. In a dc-coupled preamplifier (in Ge detectors), the *saturation behaviour is determined by its feedback resistance value R_f and its* saturation output level V_m. When saturation conditions are reached, the current flowing through R_f is simply given by V_m/R_f. This current will be supplied by the charges produced by radiation interactions in the detector. For a single pulse, the charge will be Ee/ε, where E is the average energy deposited in the detector, e is the charge on the electron and ε is the energy required to form one charge carrier pair. When multiplied by the rate at which events are occurring, a current is obtained which must equal the current flowing through R_f. Thus, if r_m is defined as the pulse rate at saturation

$$\frac{Eer_m}{\epsilon} = \frac{V_m}{R_f}, \quad \text{or} \quad r_m = \frac{V_m \epsilon}{EeR_f}$$

Since V_m and R_f are constants for a preamplifier of given design, the product Er_m, called the *energy rate limit*, can be specified to describe the conditions under which it will go into saturation

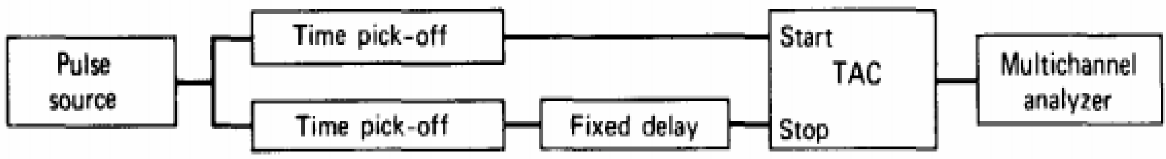
$$Er_m = \frac{V_m \epsilon}{eR_f}$$

The value of ε is appropriate to the type of detector with which the preamp is designed to operate.

(ii) Time to Amplitude Convertors (TAC)

The great popularity of TACs for the measurement of time intervals stems from the wide availability of multichannel analyzers in most measurement laboratories. By converting the time interval to a proportional pulse amplitude, the TAC allows the use of well-developed methods for the analysis and storage of pulse amplitudes as a substitute for the direct measurement of the time interval.

One of the more important properties of the TAC is the linearity of its time interval to amplitude conversion. In order to test the linearity, means must be provided for introducing fixed delays of known magnitude between start and stop pulses. For timing periods up to a few hundred nanoseconds, this can be accomplished by using variable lengths of coaxial cables. Other methods applicable for testing TAC linearity over wider time ranges are given in.



TACs are generally of two distinct types. The overlap type is based on supplying start and stop pulses of standard rectangular shape to the converter and measuring the area of overlap between the two. If the two pulses are coincident, the overlap will be complete, whereas if they are separated by a pulse width, there will be no overlap. Therefore, if an output pulse is generated whose amplitude integrates the area of overlap, the time to amplitude conversion is carried out. The principal merit of the overlap scheme is that it is very fast compared with other methods. Unfortunately, it tends to have poor linearity and accuracy specifications, and therefore it is used mainly in those applications in which maximum counting rates are of primary interest. In the start-stop type of TAC, the start pulse initiates some circuit action, such as the charging of a capacitor by a constant-current source. This action continues until terminated by the appearance of the stop pulse. The constant current generates a linear ramp voltage, which is stopped at an amplitude proportional to the interval between the start and stop pulses. Designs of this type tend to have better linearity characteristics than overlap types and are more commonly encountered in routine time spectroscopy measurements.

Q5. What is the need of preamplifier? Discuss the voltage and charge sensitive preamplifiers. Where are these preamplifiers used? (9)

Ans. A **preamplifier** (preamp or "pre") is an electronic amplifier that converts a weak electrical signal into an output signal strong enough to be noise-tolerant and

strong enough for further processing, or for sending to a power amplifier and a loudspeaker. Without this, the final signal would be noisy or distorted. They are typically used to amplify signals from analog sensors such as microphones and pickups. Because of this, the preamplifier is often placed close to the sensor to reduce the effects of noise and interference.

An ideal preamp will be linear (have a constant gain through its operating range), have high input impedance (requiring only a minimal amount of current to sense the input signal) and a low output impedance (when current is drawn from the output there is minimal change in the output voltage).

It is "a device that serves as an input selector and router between the source and a power amplifier, that can attenuate or amplify the overall volume level."

A Preamp can be a simple passive signal router with no power supply or active circuitry or it can be a complex multi-box multi-voltage device with highly sophisticated active circuitry.

Need of Preamplifier:

(i) It is used to boost the signal strength to drive the cable to the main instrument without significantly degrading the signal-to-noise ratio (SNR). The noise performance of a preamplifier is critical.

(ii) In an audio system, they are typically used to amplify signals from analog sensors to line level. The second amplifier is typically a power amplifier (power amp). The preamplifier provides voltage gain (e.g., from 10 mV to 1 V) but no significant current gain. The power amplifier provides the higher voltage necessary to drive loudspeakers. For these systems some common sensors are microphones, instrument pickups, and phonographs. Preamplifiers are often integrated into the audio inputs on mixing consoles, DJ mixers, and sound cards. They can also be stand-alone devices

(iii) It increases the energy in the detector signal to such a level that it can drive a reasonable length of low impedance coaxial cable properly terminated at the other end with only a small loss in pulse amplitude.

(iv) When the detector produces a very weak signal it may be advantageous to amplify that signal to a level where external noise becomes negligible. The weak charge signal is integrated and converted to a voltage pulse by the preamplifier without adding too much noise. The voltage signal has an amplitude that is proportional to the energy deposited in the detector.

Voltage Charge-Sensitive Configuration

Preamplifiers can be either the *voltage-sensitive* or *charge-sensitive* type. The **voltage-sensitive type** is more conventional in many electronic applications and consists simply of a configuration that provides an output pulse whose amplitude

is proportional to the amplitude of the voltage pulse supplied to its input terminals. A schematic diagram of a voltage-sensitive configuration is shown in fig. 12.

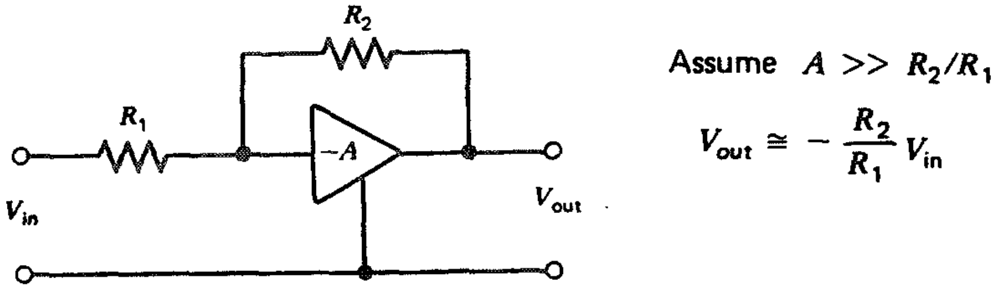


Figure 12. Schematic diagram of a simplified voltage-sensitive preamplifier configuration. R_2 is the feedback resistance.

If the time constant of the input circuit (the parallel combination of the input capacitance and resistance) is large compared with the charge collection time, then the input pulse will have an amplitude equal to

$$V_{max} = \frac{Q}{C}$$

Where C is the input capacitance. For most detectors, the input capacitance is fixed so that the output pulse produced by a voltage-sensitive preamp is proportional to the charge Q liberated by the incident radiation. If the input capacitance were to change, this desirable constant proportionality would no longer hold. For example, in semiconductor diode detectors, the detector capacitance may change with operating parameters. In these situations, a voltage-sensitive preamp is undesirable because the fixed relationship between V_{max} and Q is lost.

Current Charge-Sensitive Configuration

The elements of a charge-sensitive configuration can remedy this situation (fig. 13). For this circuit, the output voltage is proportional to the total integrated charge in the pulse provided to the input terminals, as long as the duration of the input pulse is short compared with the time constant R_fC_f. Changes in the input capacitance no longer have an appreciable effect on the output voltage. Although originally developed for use with semiconductor diode detectors, this charge-sensitive configuration has proved its superiority in a number of other applications, so that preamplifiers used with the detectors in which capacirance does not necessarily change are also often of the charge-sensitive design.

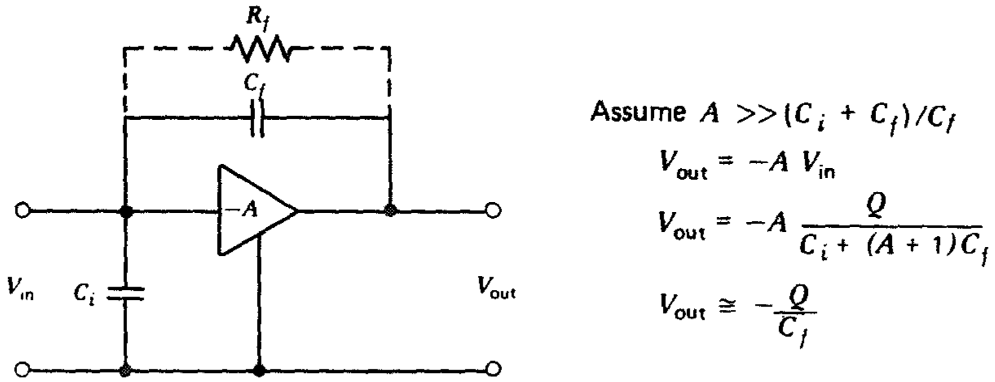


Figure 13. Simplified diagram of a charge-sensitive preamplifier configuration. If the conditions indicated are met, the output pulse amplitude becomes independent of the input capacitance C_i. The time constant given by the product $R_f C_f$ determines the decay rate of the tail of the output pulse.

Ideally, the rise time of the pulse produced from the preamplifier is determined only by the charge collection time in the detector and is independent of the capacitance of the detector preamp input. That is generally the case, but for fast detectors the rise time may also be influenced by time constants arising from several factors. Prominent among these is any series resistance from undepleted regions in semiconductor detectors or imperfect electrical contact to the detector active volume. This resistance couples with capacitance at the input of the preamplifier to define a time constant that can show the rise of the pulse if it is comparable with or exceeds the charge collection time.

Uses of Preamplifiers:

(i) They are used in signal processing for solid-state diode detectors and gas-filled ionization or proportional counters.

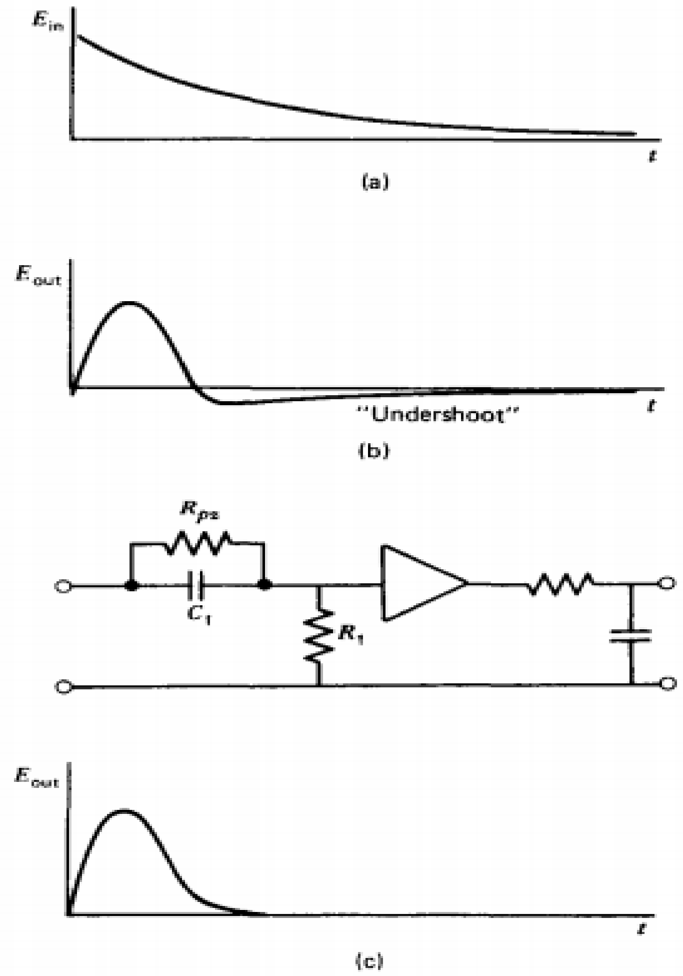
(ii) They are used in scintillator counters to avoid changes in time constant in the anode circuit of the Photomultiplier tube.

Q6. Discuss the terms pole zero cancellation and pile-up. (3)

Ans. Pole Zero Cancellation

In pulse shaping have assumed that the input pulse from the preamplifier consists of a step voltage. Although the decay of the preamplifier pulse is usually long, it is not infinite and the finite decay will have a measurable effect on the response of the networks. For example, a basic CR-RC differentiator-integrator will no longer produce a strictly unipolar response if the input pulse has the finite decay shown in fig . Instead, there will be a slight zero crossover or undershoot of the pulse, which then recovers back to zero with a time characteristic of the preamplifier decay time. Because preamplifiers have long decays (of the order of 50 ps), the undershoot persists for a relatively long time. If another pulse arrives during this

period of time, it will be superimposed on the undershoot and an error will be induced in its amplitude. The problem is particularly severe for very large signal pulses that overload the amplifier and consequently lead to rather large following undershoots.
The term pole-zero cancellation describes a technique in which the network is modified to again restore the simple exponential output without undershoot. A resistance R_{pz} is added in parallel with the capacitor of the CR network, resulting in a modified transfer function.Most spectroscopy amplifiers incorporate a pole-zero cancellation circuit to eliminate this undershoot. The benefit of pole-zero cancellation is improved peak shapes and resolution in the energy spectrum at high counting rates.



Pile-up

The shape of the pulse from the preamplifier has a quick rise time and large delay time. Due to this, a long tail appears in the output. At high rates, the subsequent pulses appear before the previous pulse falls to the baseline. This is called pile-up (fig. 14) and causes the system to saturate. Thus, the vital amplitude information is lost. It cannot be removed, but can be reduced to a certain extent. Reducing the value of the feedback resistance in the circuit can help, but only at the expense of increased noise.

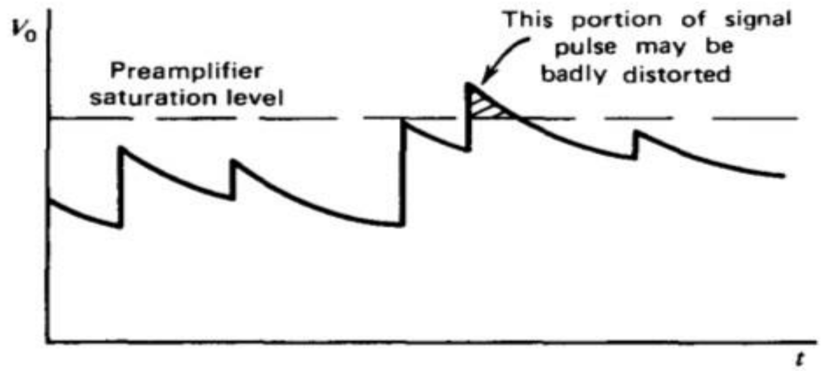
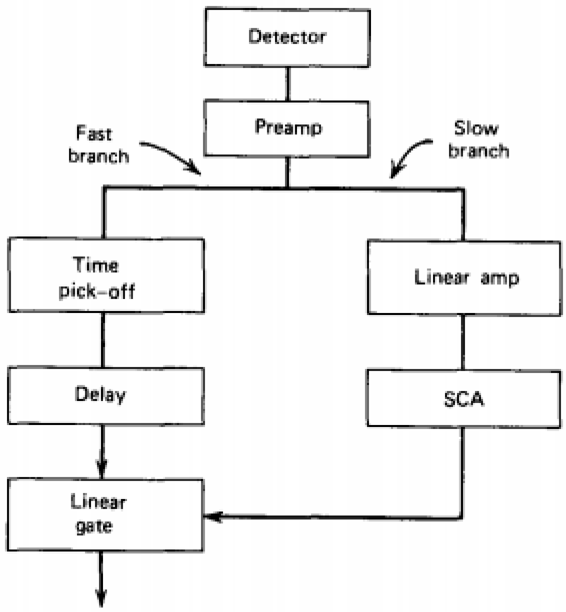


Figure 14.

Q7. Draw the diagram of gamma-gamma slow-fast coincidence set-up and discuss its working. (6)

Ans. A coincidence circuit is an electronic device with one output and two (or more) inputs. The output activates only when the circuit receives signals within a time window accepted as *at the same time* and in parallel at both inputs.



Coincidence measurements are an important tool in the detection of ionizing radiation for a wide range of applications. Many nuclear processes produce two photons simultaneously, while other processes produce two or more photons in quick succession. In such cases, it is possible to study the temporal and angular correlations between the two photons by setting up a coincidence detector system. These emissions can occur simultaneously or within a time period that is very short compared to the time resolution of the detection system. For example, decay by beta emission to a daughter nucleus, which in turn decays by gamma emission, produces the beta particle and the gamma ray at essentially the same time. Similarly, one nucleus may emit several gamma rays in cascade, which are effectively simultaneous because the delay between the events is short. Delays of 10^{-9} seconds are common.

In nuclear physics applications, coincidence systems are used to detect and identify weak detection signals or to distinguish a physics signal from background signals, as is done in Compton suppression or cosmic veto systems. In high-energy or particle physics, detection systems consisting of thousands of detectors and electronics channels are all operated in coincidence when two accelerated beams collide to search for newly formed particles or new decay pathways.

Time-coincidence measurements

In addition to radiation detector characteristics such as efficiency or energy resolution, time resolution is important to measure. This is required to determine the time dependence of nuclear decays as discussed above. It reflects the ability to measure the arrival time of the incident particle or the time of a specific interaction and its associated signal. The time resolution of a specific detector depends on several parameters such as signal shape called “walk time” and signal noise called “jitter time”. In some cases the actual time difference between the two events can be measured, but in many cases it is only necessary to determine that the events are correlated in time. The determination that two nuclear events occur at the same time is made electronically with a coincidence system. This unit operates on standardized pulses and determines whether events occur within a certain time interval, called the resolving time. The standard pulses from any single channel analyzer are used as input, with one input from each detector.
The number of coincidences that are real, not random, is determined by the physics of the decay and the solid angles and efficiencies of the detectors. These are the true coincidences. In some experiments, the number of coincidences is the only information needed. Often, however, the coincidence signal is used to open the linear gate in a multichannel analyzer so that a spectrum is acquired under coincidence conditions. This experiment has three separate parts which use the coincidence technique in different ways.

γγ Angular correlations

The angular correlation of two gamma rays, γ_1 and γ_2 may be defined as the probability of γ_2 being emitted at an angle relative to the direction of γ_1. The emission of gamma rays from excited nuclei can be treated mathematically as the classical radiation of electromagnetic energy from a charged system. The electric field can be expanded in vector spherical harmonics, corresponding to the various multipoles of the charge distribution. The shape of the angular distribution of the radiation with respect to the radiating system is uniquely determined by the order of the multipole.
In nuclear systems, the order of the multipole depends upon the angular momentum numbers and the parities of the initial and final states involved in the transition. Thus, if all nuclei in a radioactive sample could be oriented so that their nuclear angular momenta were aligned, the shape of the angular distribution of gamma rays could be used to determine the multipolarity of the transition.

However, nuclei are randomly oriented. Very strong magnetic fields at low temperature could be used to provide orientation, but a simpler method is to use coincidence technique for cases in which two or more gamma rays are emitted in cascade. The first gamma ray establishes the direction of the spin axis of the nucleus; hence the second gamma ray will have a definite distribution with respect to this axis. One needs only to measure the angular correlation for the two gamma rays and compare it with the tabulated values for various multipolarities.

In nuclear physics experiments the angular correlation is measured between two gamma rays, which are emitted almost simultaneously in the cascade from the decay of a radioactive nucleus. The gamma rays are detected using two NaI scintillation counters in which the height of the electronic output pulses is proportional to the incident gamma ray energies. By pulse height selection, in "single channel analyzer mode", one counter will be used to record γ_1 and the other γ_2. The counting rate of each counter, R_i, for detecting its selected gamma ray is given by:

$$R_i = N_0\, \varepsilon_i\, \frac{\Omega_i}{4\pi} \qquad i = 1,2$$

where:

N_0 is the number of decays per second in the radioactive source.
e_i is the efficiency of the detector.
O_i is the geometrical solid angle subtended by the detector.
I_n the absence of the angular correlations, the true rate of detected gamma-ray coincidences is:

$$R_{true} = R_1\, \varepsilon_2\, \frac{\Omega_2}{4\pi} = N_0\, \varepsilon_1\, \frac{\Omega_1}{4\pi}\, \varepsilon_2\, \frac{\Omega_2}{4\pi}$$

The random coincidence rate between a gamma ray detected in Detector 1 and a gamma ray detected in Detector 2 is:

$$R_{random} = R_1\, R_2\, \Delta t = N_0^2\, \varepsilon_1\, \frac{\Omega_1}{4\pi}\, \varepsilon_2\, \frac{\Omega_2}{4\pi}\, \Delta t$$

where Δt is the resolving time of the coincidence counting system between the two detectors. If both detectors are set to respond to both gamma rays, the counting rate in each detector is the sum of counting rates for different gamma rays.

Q8. Draw block diagram of Multichannel analyzer and discuss its working in Pulse height analysis mode. (4)

Ans. The Multichannel analyzer (MCA) is comprised of the components as shown in fig. 15. Its operation is based on the principal of converting an analog signal (the pulse amplitude) to an equivalent digital number. Once this conversion has been done, the available technology is used for the storage and display of digital information and solve the problem of recording the pulse height spectra.

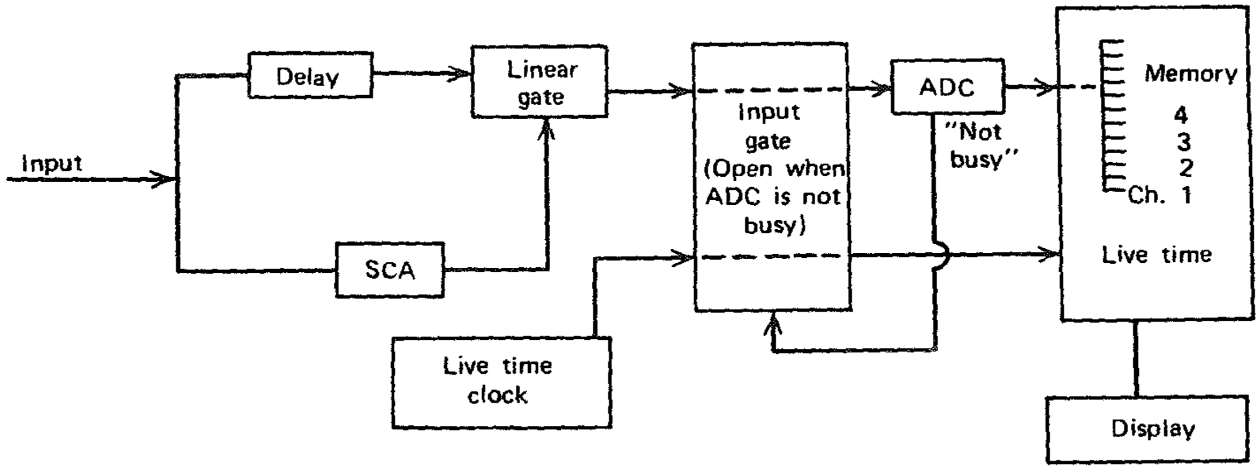


Figure 15. Functional block diagram of a typical MCA.

In pulse-height analysis (PHA) mode, *the pulses are counted based on their amplitude.* The MCA works by digitizing the amplitude of the incoming pulse with an analog-to-digital convertor (ADC). The MCA then takes this number and increments a memory channel whose address is proportional to the digitized value. In this way incoming pulses are sorted out according to pulses height and the number at each pulse height stored in memory locations corresponding to these amplitudes. The total number of channels into which the voltage range is digitized is known as *conversion gain.* This determines the resolution of the MCA.
The heart of the MCA is the ADC and in order to give it sufficient time to digitize the input signals, there are usually very strong requirements on the rise times and widths of the incoming analog signals. To facilitate this, there exist a number of modules used mainly for adapting signals for MCAs. These include the biased amplifier and the pulse stretcher.

Q9. Discuss RC-CR and RC-(CR)^n methods of wave shaping. **(6)**

Ans. <u>**RC-CR SHAPING**</u>

The output of a single differentiating network (fig. 16) has a sharp pointed top which makes subsequent pulses height analysis difficult because the maximum pulse amplitude is maintained only for a very short time period. Furthermore, because the differentiation allows all high-frequency components of any noise mixed with the signal to be passed by the network, the signal-to-noise characteristics of the network in practical applications are usually very poor.

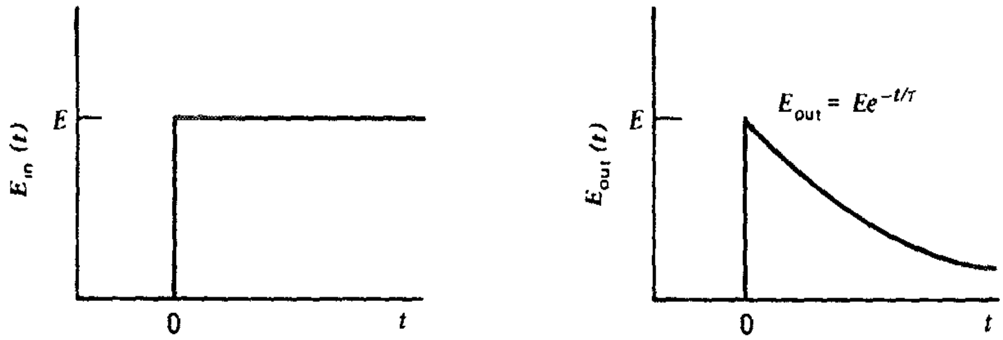


Figure 16

But, if a stage of RC integration is added following differentiation, both of these drawbacks are considerably improved. The combination of a single stage of differentiation followed by a single stage of integration is in fact a common method of shaping preamplifier pulses.
The fig. 17 shows the basic CR-RC shaping network. An ideal unity-gain operational amplifier (with infinite input impedance and zero output impedance) separates the two individual networks for impedance isolation so that neither network influences the operation of the other.

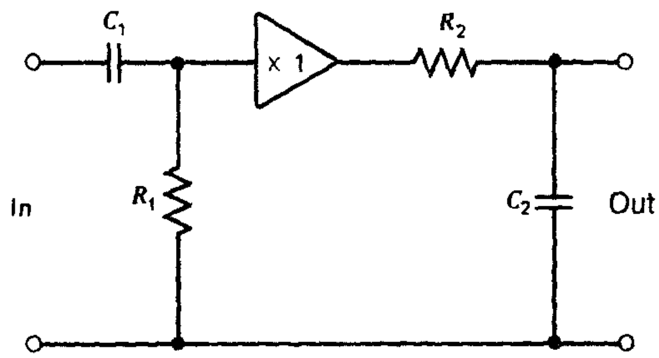


Figure 17 A shaping network consisting of sequential differentiating and integrating stages, sometimes denoted a *CR-RC* network.

The general solution of the response of the combined network to a step voltage of amplitude E at t=0 is ...(1)

$$E_{\text{out}} = \frac{E\tau_1}{\tau_1 - \tau_2}\left(e^{-t/\tau_1} - e^{-t/\tau_2}\right)$$

Where τ_1 and τ_2 are time constants of the differentiating and integrating networks, respectively. Plots of this response for several different combinations are shown in fig. 18. In nuclear amplifiers, CR-RC shaping is most often carried out using equal differentiation and integration time constants. In this case, eq (1) becomes indeterminant, and a particular solution for this case is

$$E_{\text{out}} = E\,\frac{t}{\tau}\,e^{-t/\tau}$$

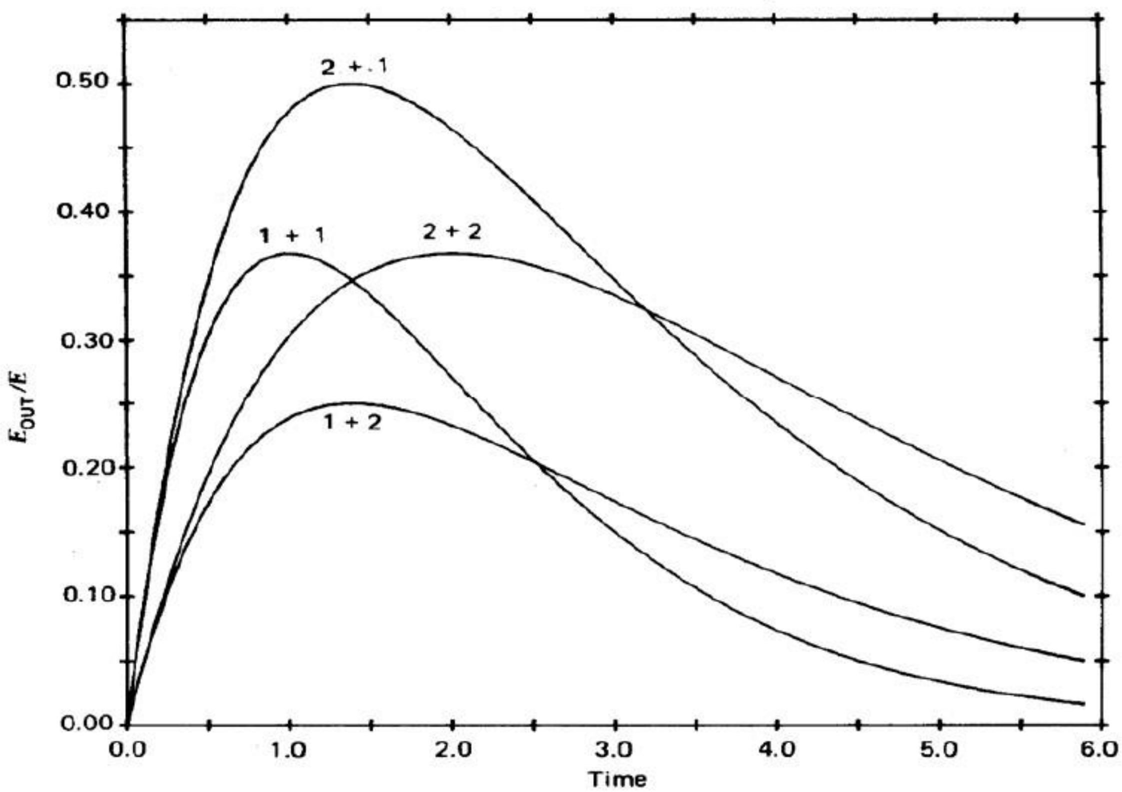


Figure 18

The choice of shaping time constants is determined primarily by the charge collection time in the detector. As always, there are competing factors that need to be considered:

- In order to minimize pile-up effects, the time constants should be kept **short** so that the pulse returns to baseline as quickly as possible.
- On the other hand, when the time constants become comparable to the rise time of the pulse from the preamplifier the input to the network no longer looks like a step voltage and the result is that some of the amplitude of the signal is lost.

This is called **ballistic deficit** and can be avoided only by keeping the time constants **long** compared to the charge collection time of the detector.
Typical values for τrange from a few tenths of a μs for 'small' semiconductor detectors through to a few μs for 'large' Ge detectors to tens of μs for some types of proportional counter.

RC-(CR)^n OR GAUSSIAN SHAPING

If a single CR differentiation is followed by several stages of RC integration, a pulse shape that approaches a Gaussian shape is realized. If the differentiation and n integration time constants are all the same value τ, the particular solution of the corresponding equation is

$$E_{\text{out}} = E\left(\frac{t}{\tau}\right)^n e^{-t/\tau}$$

In practice, four stages of integration (n=4) are sufficient so that the difference between the resulting pulse shape and a true Gaussian is negligible. The time

required for the shaped pulse to reach its maximum amplitude (called the *peaking time*) is equal to nτ.

Four equal time constants throughout a RC-(CR)^4 network results in a peaking time that is a factor of 4 longer than that for a simple CR-RC network. However, if the time constants are adjusted to give equal peaking times for the two methods, the more symmetric shape of the Gaussian pulse results in a faster return to the baseline. Pulse pile-up at high counting rates is thereby reduced. Gaussian shaping also has the advantage of better signal-to-noise characteristics for individual pulses compared with CR-RC shaping (fig. 19).

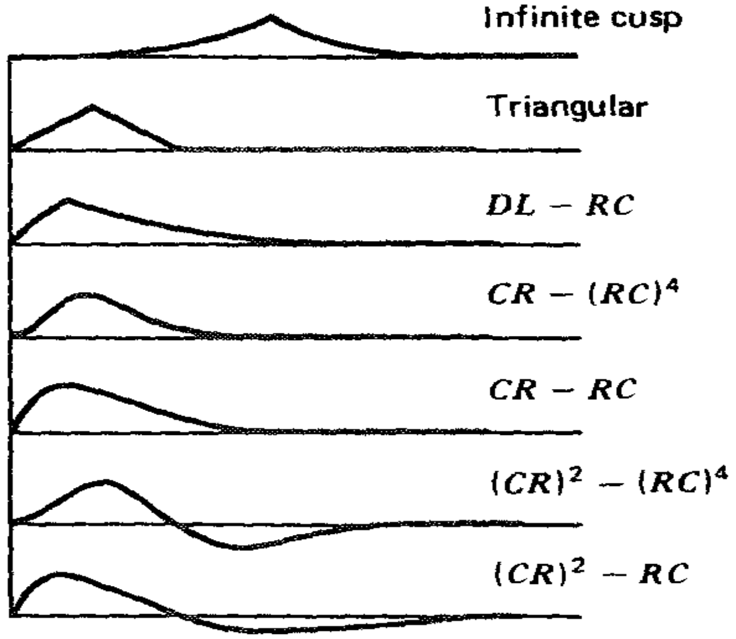


Figure 19. Various pulse shapes and signal-to-noise ratio relative to the infinite cusp.